

25.4 Active 2nd-Order Intermodulation Calibration for Direct-Conversion Receivers

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In direct-conversion receivers, the second-order intermodulation (IM2) distortion generated in the RF front-end may severely degrade the system SNR if it is not controlled below a certain level. Usually the down-conversion mixer is the dominant IM2 source in the RF front-end because of ac coupling between the LNA and mixer has filtered out most of the LNA's contribution. There are some mixer topologies [1, 2] that can reduce this nonlinear effect, but they may not be applicable to suppress IM2 distortion sufficiently for some systems because of conflicting mixer requirements, on-chip LO mismatch and limited die area. To resolve this problem, several IM2 calibration methods were presented in [3, 4], but are less attractive because of small bandwidth or limited IM2 cancellation capability. In this paper we introduce an active IM2 calibration circuit that works with an International Mobile Telecommunications (IMT-2100 WCDMA) direct-conversion mixer without the abovementioned issues.

If two arbitrary tones ($V_1 \cos(\omega_1 t + \theta_1)$ and $V_2 \cos(\omega_2 t + \theta_2)$) are present at the mixer RF port and $f(\omega_o t + \theta_o)$ is the signal at the LO port, the IM2 distortion current at the mixer output can be represented as $A_1 \cos[(\omega_1 - \omega_2)t + (\theta_1 - \theta_2)]$. Similarly if these two tones are fed into a nonlinear IM2 generator, its output current can be represented as $A_2 \cos[(\omega_1 - \omega_2)t + (\theta_1 - \theta_2)]$. When these two currents are summed together, the resulting IM2 distortion can be reduced dramatically if A_2 is adjusted to be $-A_1$. Generally, when multiple tones are present in the RF spectrum, numerous beat frequencies are formed at the mixer output. Meanwhile, the IM2 generator also produces beats with the same frequencies and initial phases as those from the mixer. Furthermore, their intermodulation coefficients are beat-frequency independent, therefore a single two-tone IM2 calibration can suffice to cancel all IM2 components satisfactorily.

This calibration idea is implemented with an IMT-2100 mixer (2110 to 2170MHz), which is shown in Figure 25.4.1. The IM2 generator produces the IM2 calibration currents while suppressing the signal current. These calibration currents are subsequently fed into two scaling units, which adjust the currents' polarities and magnitudes [4]. At the I/Q mixer outputs, the original mixer IM2 currents are combined with the calibration IM2 currents and the IM2 distortion terms cancel.

The schematic of the IM2 generator is shown in Figure 25.4.2. Differential RF signals are applied to $M1$ and $M2$, which have their drain currents summed together. This is a squaring circuit and it generates enough IM2 current to cancel the mixer IM2 distortion even under the worst mismatch scenario. To avoid excessive loading of the mixer, the dimensions of transistor $M1$ and $M2$ are chosen to be small. Transistors $M4$ to $M11$ duplicate and split the calibration current, thereby creating differential inputs to the I and Q scaling units. To further reduce the loading of the mixer and to reduce the noise contribution, two RF amplifiers are added in Figure 25.4.2. Figure 25.4.3 shows the schematic for these amplifiers, including the bias circuit. Let $-A_v$ be the voltage gain for each amplifier, the IM2 calibration current is then boosted by $(1 + A_v)^2$. This gain-boosting technique also allows us to reduce the widths of $M1$ and $M2$ by $(1 + A_v)^2$ to save dc current consumption and minimize the thermal noise caused by large transconductance (g_m). Ideally, the gains of the RF amplifiers should be constant across the full bandwidth from 2110 to 2170MHz or the calibration performance will vary with frequency.

Another benefit of using these RF amplifiers is that their gain variations over temperature can be adjusted to compensate for the temperature variation of the mixer IM2 distortion. This adjustment can be achieved by varying the temperature coefficient of the bias resistor R as shown in Figure 25.4.3. Assuming a quadratic I - V law, the bias current and mirrored RF-amplifier dc currents, are proportional to R^{-2} . In reality, short channel effects alter this relationship to be $R^{-1.82}$. Hence by adjusting R 's temperature coefficient (TC), we can make the bias current change with temperature so that the generated IM2 current varies in the same way as the mixer IM2 current to achieve proper cancellation over temperature. A practical way to implement this resistor is shown in Figure 25.4.4, in which one resistor branch is selected by three control bits to cope with the unknown mixer IM2 temperature variations and resistor process variations. A fourth bit is used to select the amplifiers' gains (high or low) to fine-tune the calibration characteristics. The IM2 calibration current is then fed into the scaling units, which, under the control of 6b calibration code x , modify the magnitude and polarity of the output differential IM2 calibration currents according to $I_{dm}^{out} = (x - 31.5)/31.5 * I_{dm}^{in}$ ($0 \leq x \leq 63$).

The calibration circuit is fabricated in a 0.25 μ m CMOS process with the IMT-2100 direct-conversion mixer. Calibration characteristics are tested under various frequency and temperature settings. The mixer is driven by an on-chip VCO, therefore the mismatch in the LO path has also been included. Figure 25.4.5 displays the measurement results for multiple channel frequencies. The horizontal axis is the 6b calibration code x , while the vertical axis is the carrier-to-IM2-distortion ratio in dB at the mixer output; the system requirement is 15dB. Note that the carrier-to-IM2-distortion ratio can be improved by more than 20dB with this IM2 calibration. Although the calibration profile shifts when frequency is swept from high to low, we can easily choose a calibration code, such as 21 to 23, that meets the system specification. Figure 25.4.6 displays the measurement results when temperature is swept from -30 to $+85^\circ\text{C}$. By choosing $x = 21$, both cold and hot IM2 requirements can be met. It is believed that lower temperature variation can be attained by further optimizing the bias resistor TC. Unlike the calibration technique in [3], the larger bandwidth of the IM2 generator also makes this approach more suitable for wideband modulation applications, such as WCDMA. The circuit micrograph is shown in Figure 25.4.7. The calibration circuit consumes 1.5mA and the noise contribution to the mixer is 0.1dB.

References:

- [1] E.E. Bautista et al., "A High IIP2 Downconversion Mixer Using Dynamic Matching," *IEEE J. Solid-State Circuits*, Vol. 35, No. 12, pp. 1934-1941, Dec., 2000.
- [2] M. Brandolini et al., "A CMOS Direct Down-Converter With +78dBm Minimum IIP2 for 3G Cell-Phones," *ISSCC Dig. Tech. Papers*, pp. 320-321, Feb., 2005.
- [3] K. Kivekas et al., "Calibration Techniques of Active BiCMOS Mixers," *IEEE J. Solid-State Circuits*, Vol. 37, No. 6, pp. 766-769, June, 2002.
- [4] P.J. Shah, "Distortion Reduction Calibration." US patent application 20030186664.

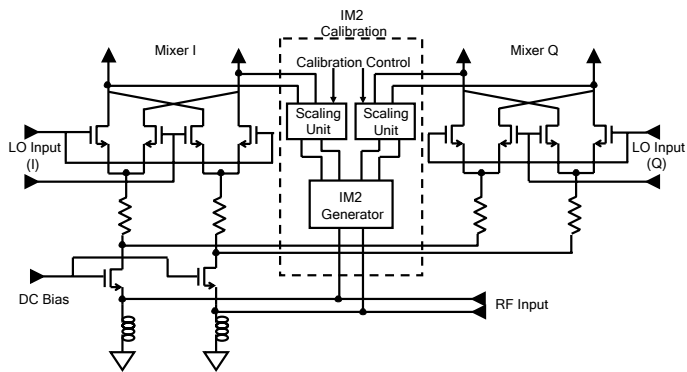


Figure 25.4.1: IM2 calibration system.

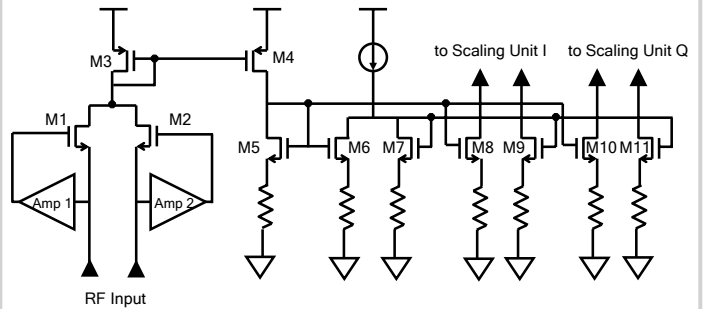


Figure 25.4.2: IM2 generator.

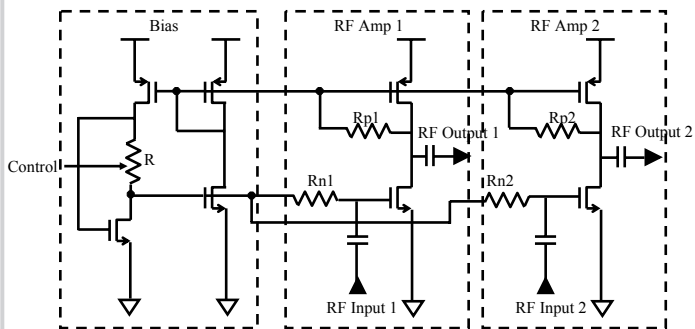


Figure 25.4.3: RF amplifiers with bias circuit.

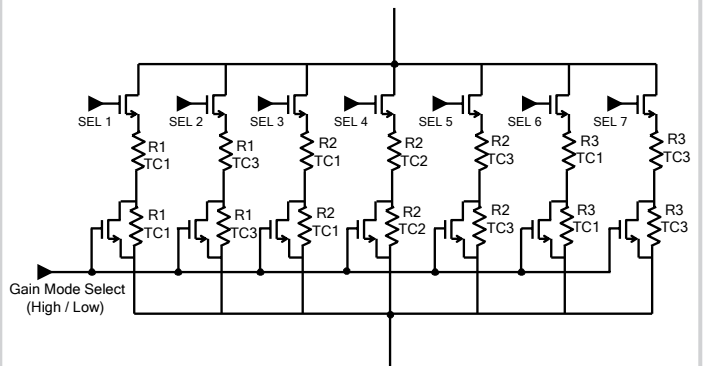


Figure 25.4.4: Bias resistor array implementation.

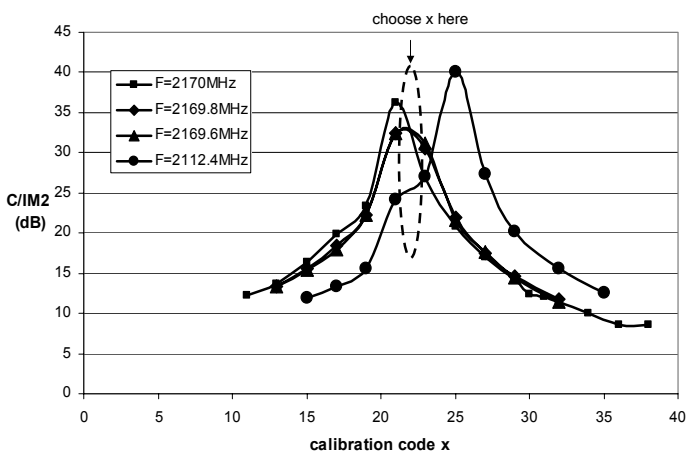


Figure 25.4.5: Measured IM2 calibration versus frequency.

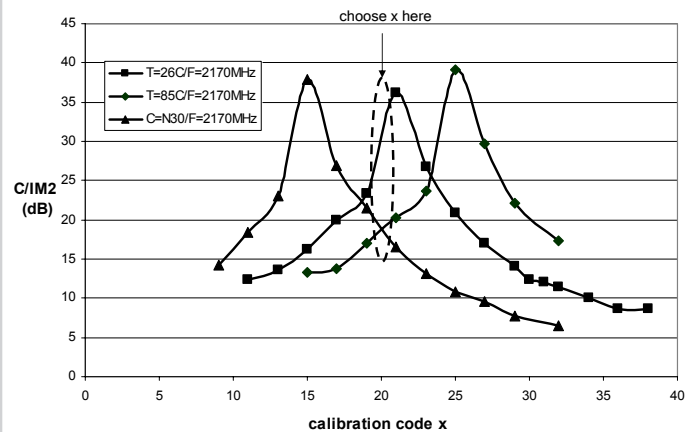


Figure 25.4.6: Measured IM2 calibration versus temperature.

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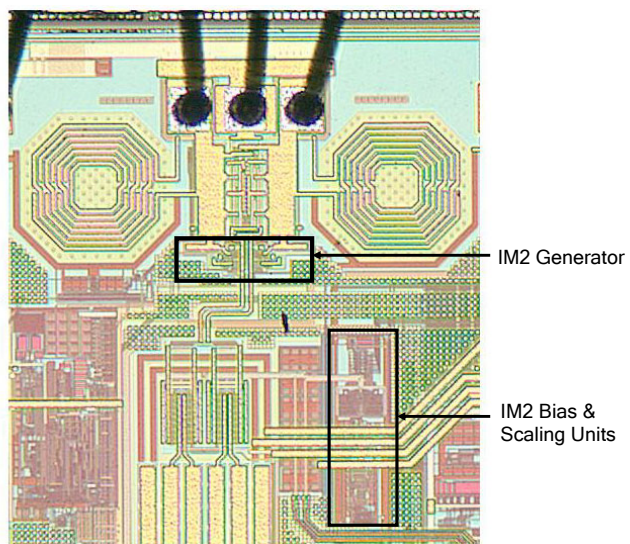


Figure 25.4.7: Micrograph of the IM2 calibration mixer.